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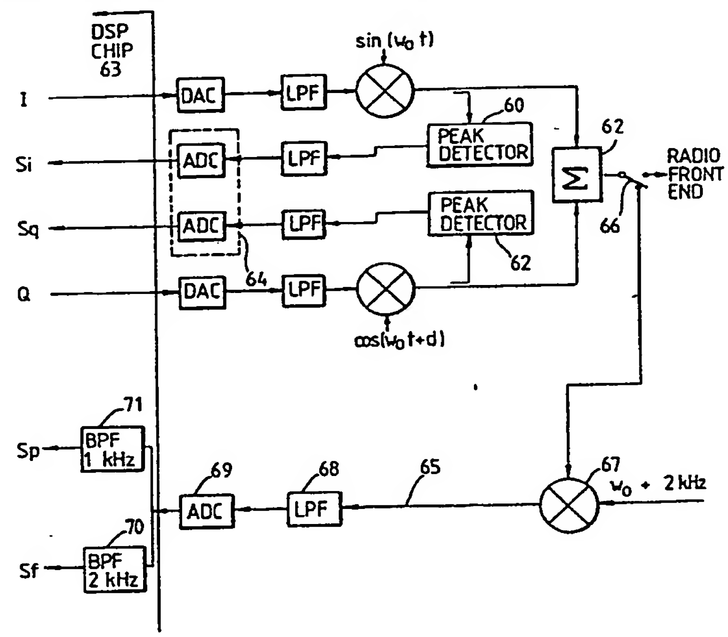
(54) **Zero-IF transmitter with error correction**

(57) A method for the correction of errors in a zero-IF transmitter using its SSB mode resides in the steps of successively a) reducing local oscillator (LO) feedthrough, b) effecting balancing of the amplitudes of quadrature channels I, Q and c) reducing the phase deviation from quadrature in the channels. The method requires the introduction of an additional signal path in the radio, using an auxiliary LO the frequency of which is offset from the rf oscillator frequency by a small amount, eg 2kHz, and an extra mixer 67 to mix, in a calibration mode, the transmitter output with the auxiliary oscillator signal to form baseband error signals $S_i$,$S_p$ for feedback 65 to the digital signal processor 63. Recursive algorithms make adjustments so that a) a first error signal $S_i$ indicative of LO feedthrough is minimised; b) the ratio Si/Sq of peak detector outputs 60,62 is then adjusted to achieve channel balance; and c) a second error signal Sp indicative of deviation from phase quadrature is finally reduced to zero. The calibration is repeated for various LO frequencies and each result stored in RAM.

Fig.6




At least one drawing originally filed was informal and the print reproduced here is taken from a later filed formal copy.

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2213006

Fig.1

PRIOR ART

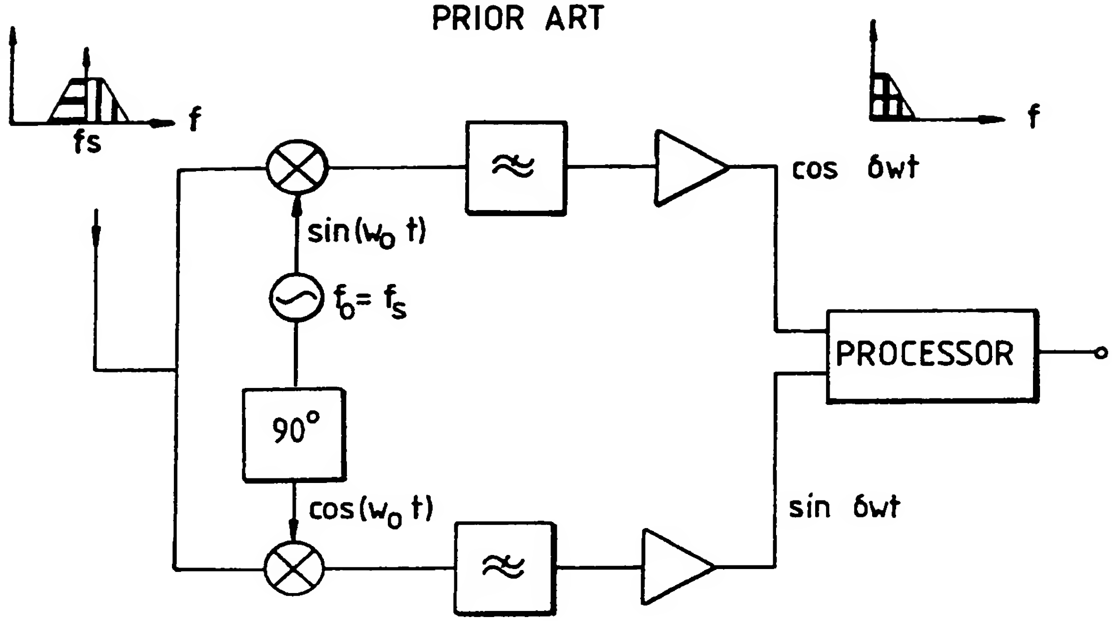



Fig.2

PRIOR ART

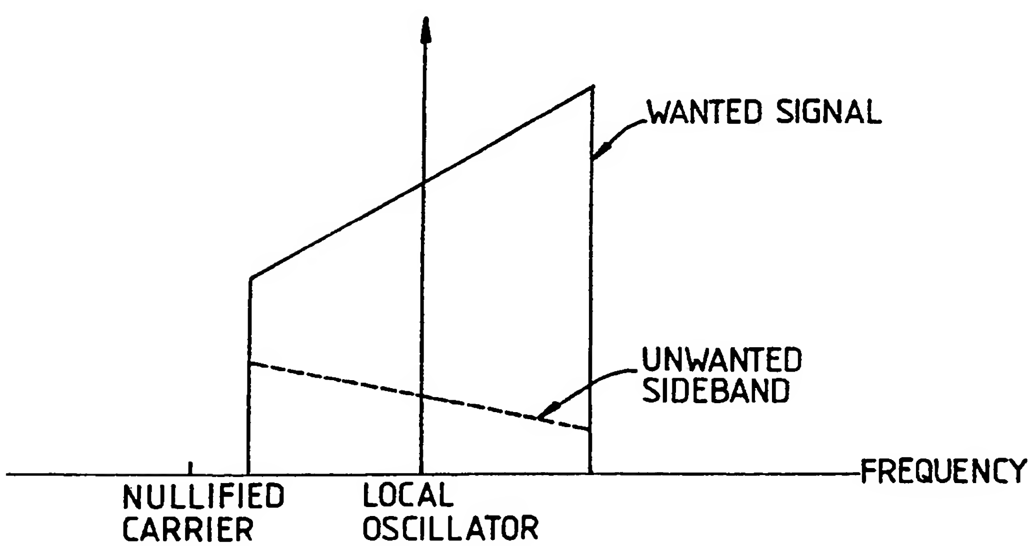


Fig. 3

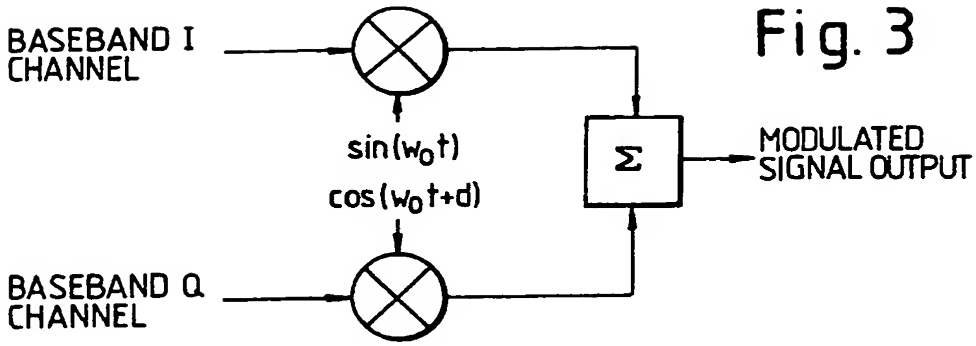

BASEBAND I CHANNEL
sin(w0t)
cos(w0t+d)
BASEBAND Q CHANNEL
Σ
MODULATED SIGNAL OUTPUT

Fig. 4

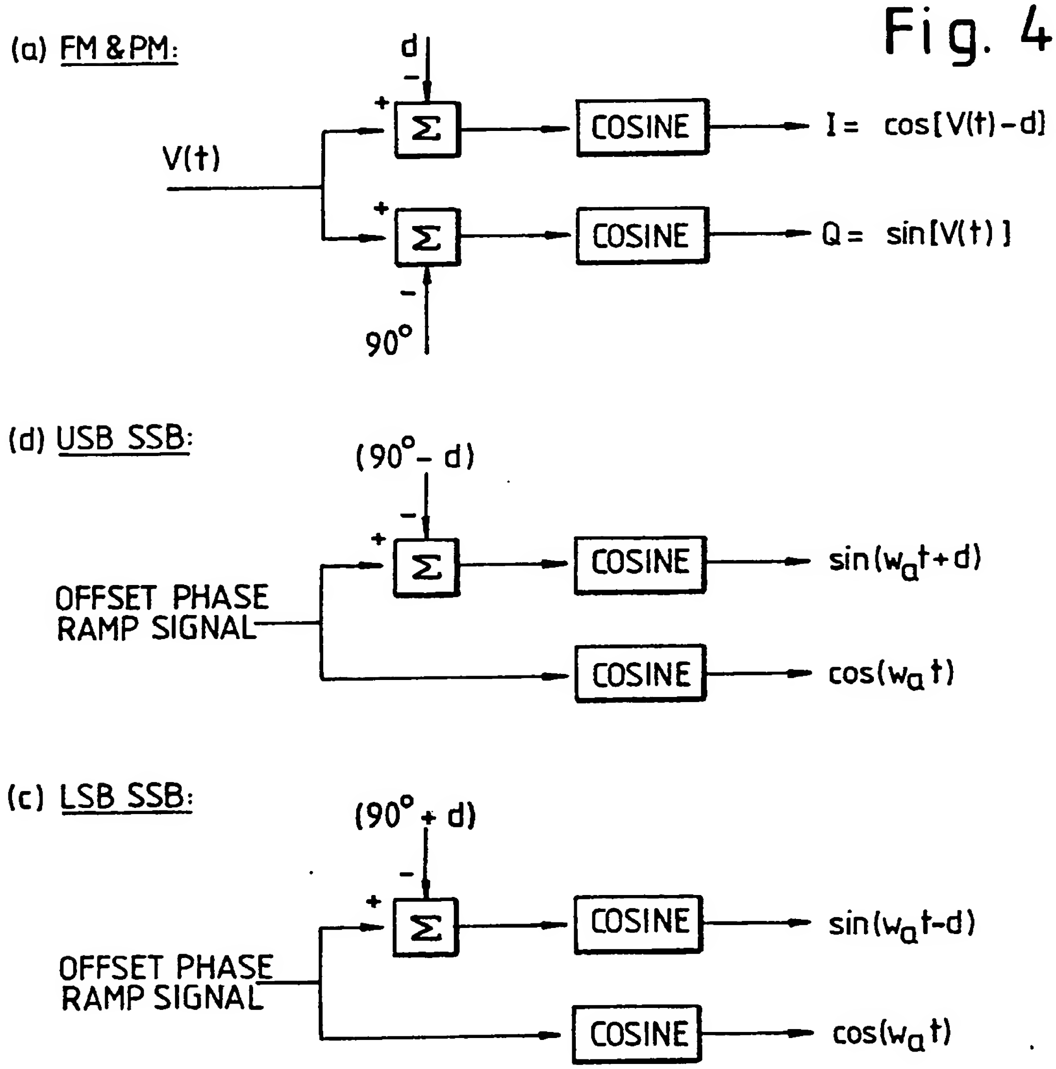

(a) FM & PM:
d
+
−
Σ
V(t)
COSINE
I = cos[V(t)−d]
+
Σ
−
90°
COSINE
Q = sin[V(t)]
(d) USB SSB:
(90°− d)
+
−
Σ
OFFSET PHASE RAMP SIGNAL
COSINE
sin(wat+d)
COSINE
cos(wat)
(c) LSB SSB:
(90° + d)
+
−
Σ
OFFSET PHASE RAMP SIGNAL
COSINE
sin(wat-d)
COSINE
cos(wat)

Fig.5

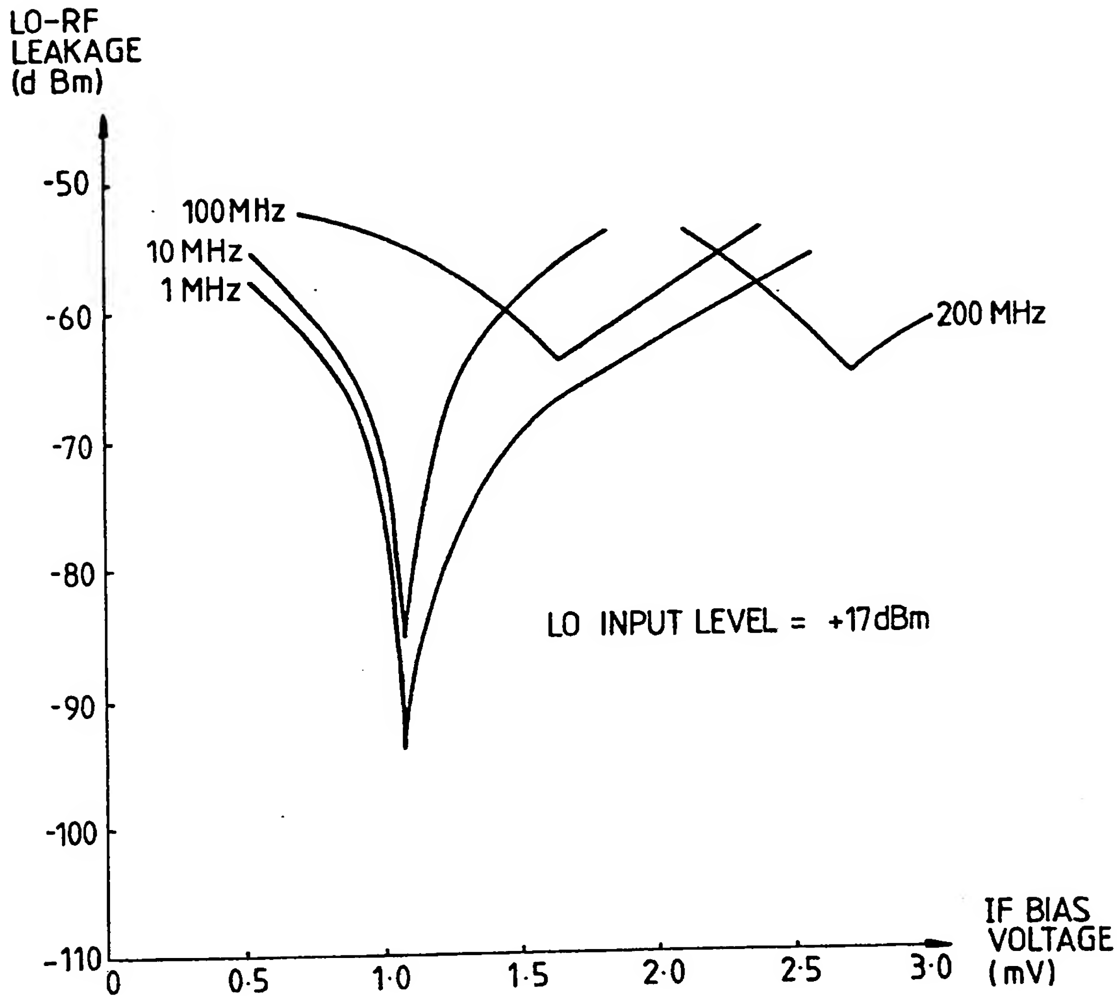

LO-RF
LEAKAGE
(d Bm)
-50
-60
-70
-80
-90
-100
-110
100MHz
10 MHz
1 MHz
200 MHz
LO INPUT LEVEL = +17dBm
0
0·5
1·0
1·5
2·0
2·5
3·0
IF BIAS
VOLTAGE
(mV)

Fig. 6

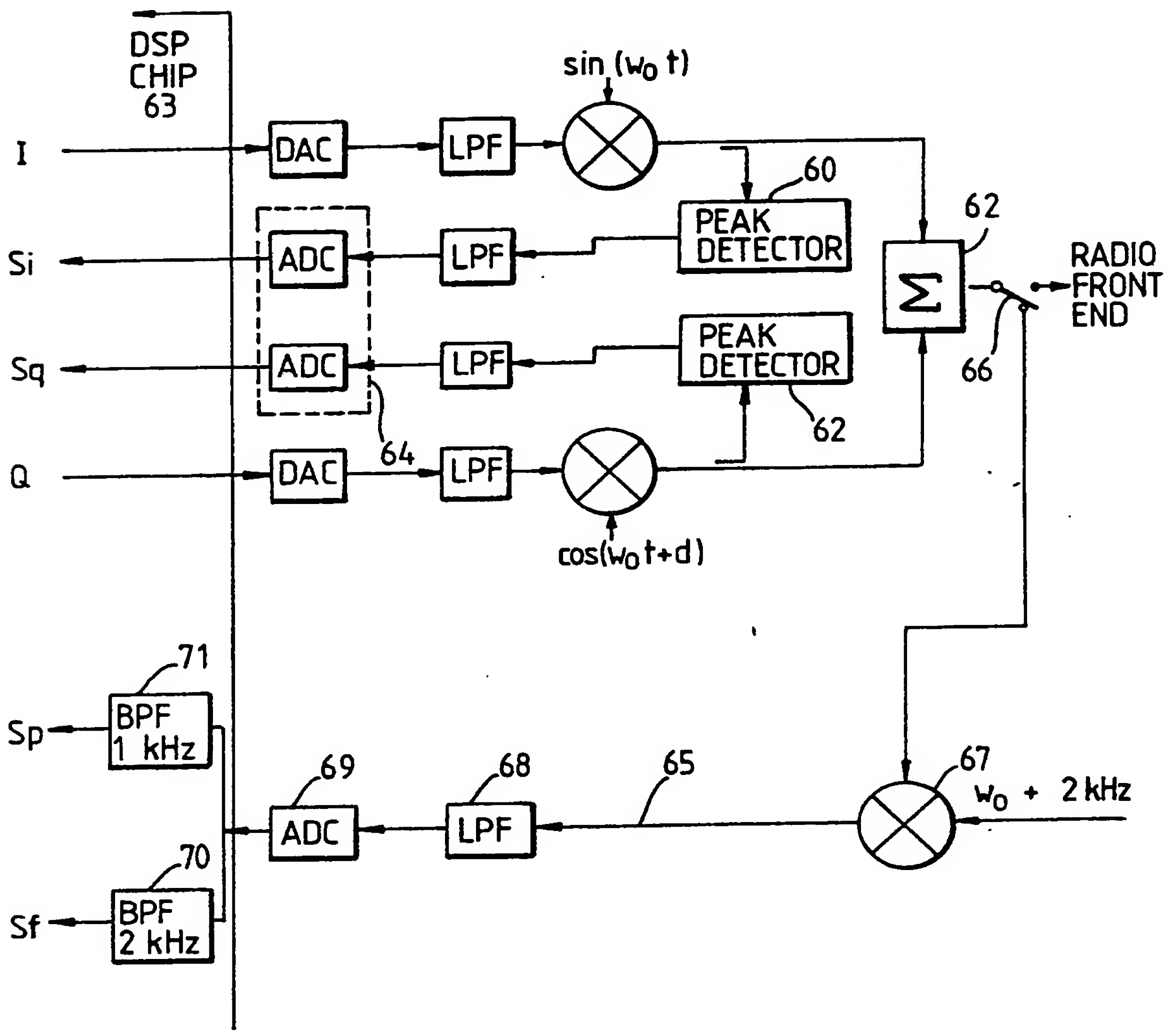

DSP
CHIP
63
I
Si
Sq
Q
DAC
ADC
ADC
DAC
LPF
LPF
LPF
LPF
sin (wo t)
cos(wo t+d)
PEAK
DETECTOR
PEAK
DETECTOR
60
62
62
64
66
RADIO
FRONT
END
71
70
Sp
Sf
BPF
1 kHz
BPF
2 kHz
69
68
65
67
ADC
LPF
wo + 2 kHz

Fig. 7

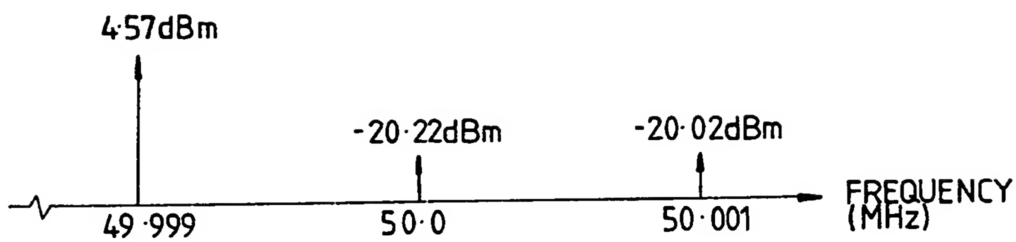

4·57dBm
-20·22dBm
-20·02dBm
FREQUENCY (MHz)
49·999
50·0
50·001

Fig. 8

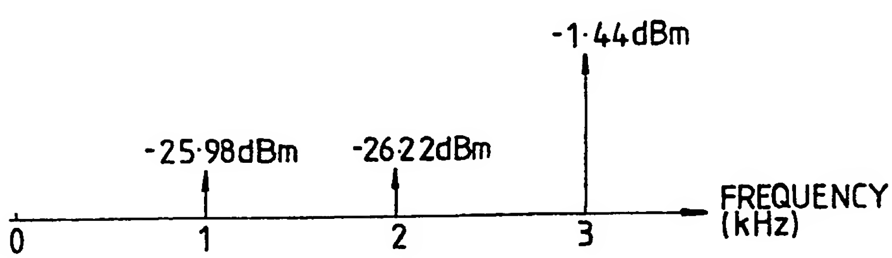
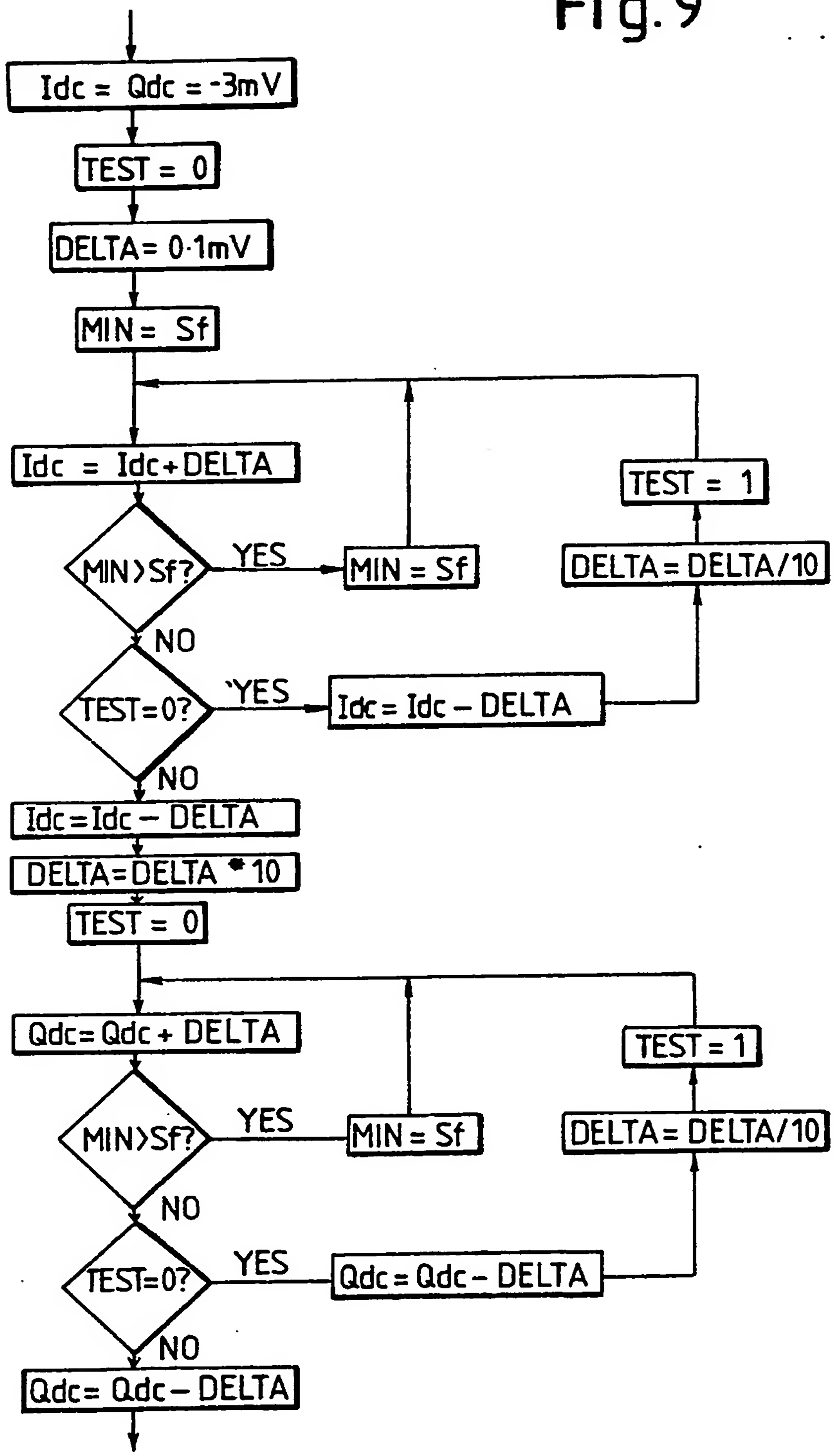
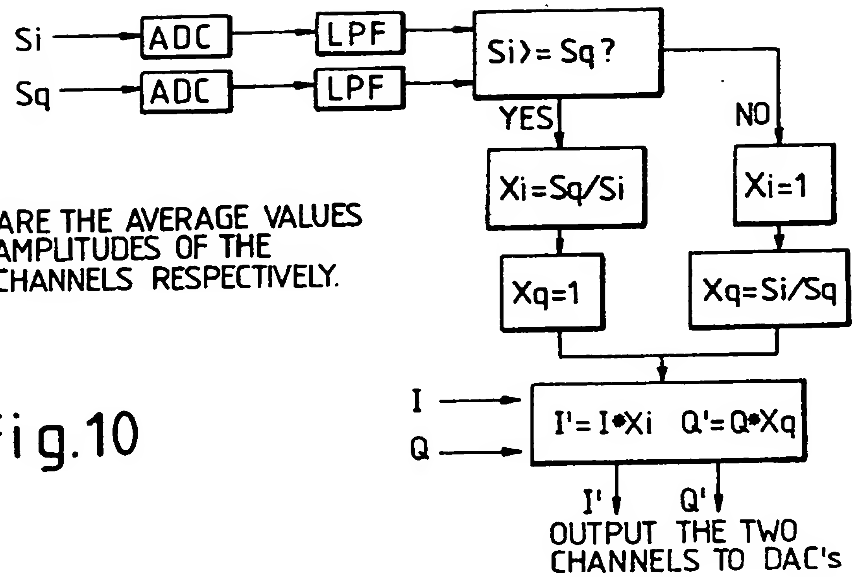

-1·44dBm
-25·98dBm
-26·22dBm
FREQUENCY (kHz)
0
1
2
3

Fig. 9


Idc = Qdc = -3mV
TEST = 0
DELTA = 0·1mV
MIN = Sf
Idc = Idc+DELTA
MIN>Sf?
YES
MIN = Sf
NO
TEST=0?
YES
Idc = Idc – DELTA
DELTA = DELTA/10
TEST = 1
NO
Idc = Idc – DELTA
DELTA = DELTA * 10
TEST = 0
Qdc = Qdc + DELTA
MIN>Sf?
YES
MIN = Sf
NO
TEST=0?
YES
Qdc = Qdc – DELTA
DELTA = DELTA/10
TEST = 1
NO
Qdc = Qdc – DELTA




Si AND Sq ARE THE AVERAGE VALUES OF THE RF AMPLITUDES OF THE I AND Q CHANNELS RESPECTIVELY.

Fig.10

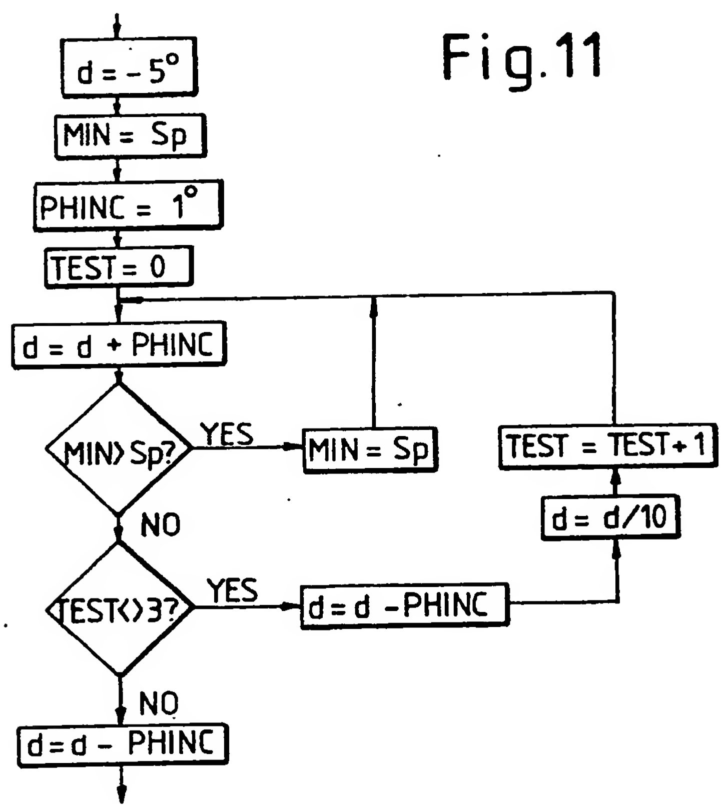



Fig.11

Fig.12

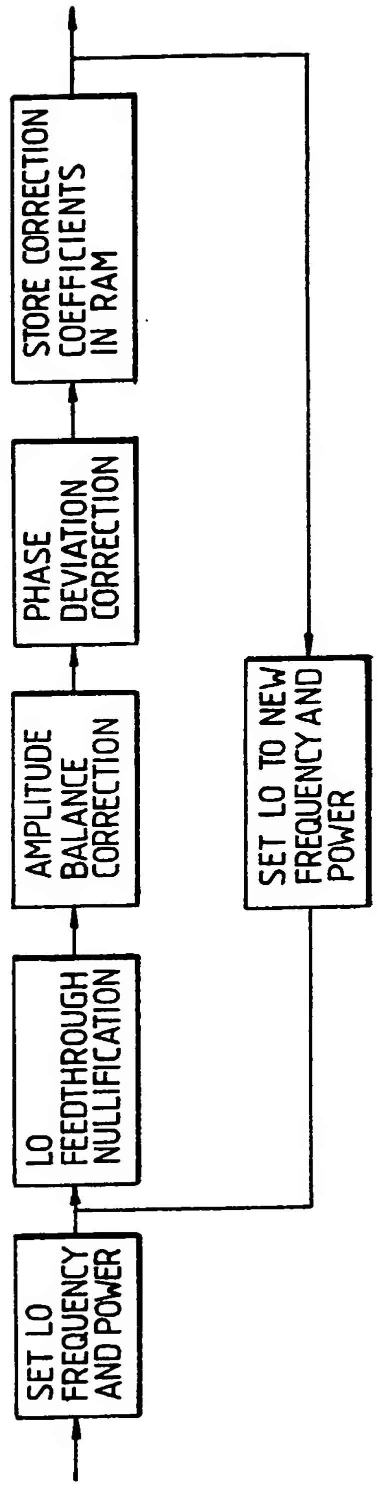

SET LO FREQUENCY AND POWER
LO FEEDTHROUGH NULLIFICATION
AMPLITUDE BALANCE CORRECTION
PHASE DEVIATION CORRECTION
STORE CORRECTION COEFFICIENTS IN RAM
SET LO TO NEW FREQUENCY AND POWER

Fig.13

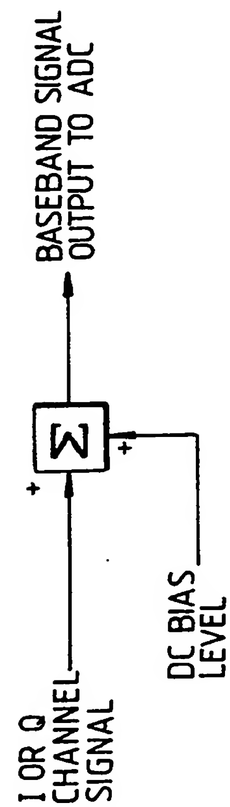

I OR Q CHANNEL SIGNAL
DC BIAS LEVEL
+
+
Σ
BASEBAND SIGNAL OUTPUT TO ADC

Fig.14

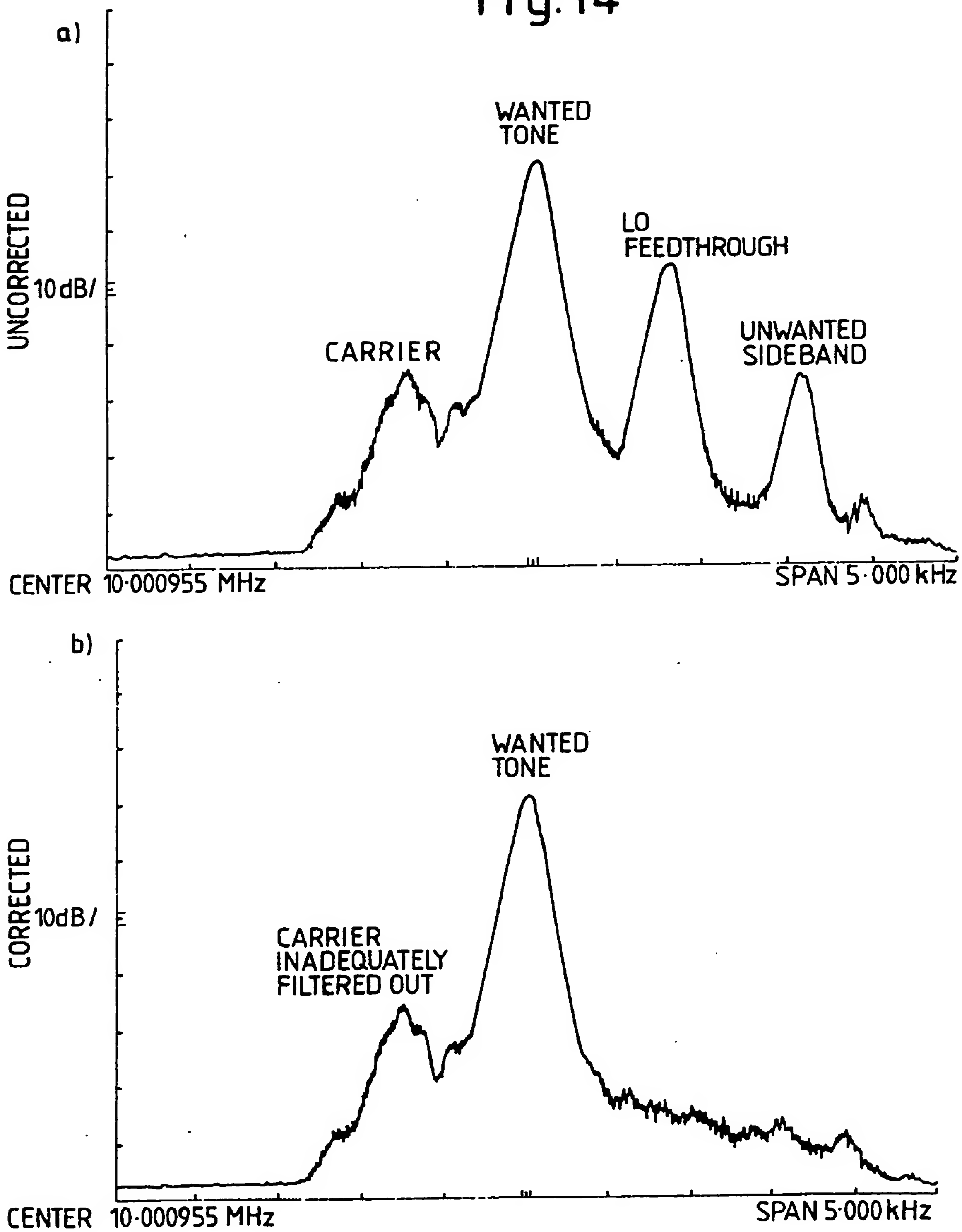

a)
UNCORRECTED
10dB/
WANTED TONE
LO FEEDTHROUGH
CARRIER
UNWANTED SIDEBAND
CENTER 10·000955 MHz
SPAN 5·000 kHz
b)
CORRECTED
10dB/
WANTED TONE
CARRIER INADEQUATELY FILTERED OUT
CENTER 10·000955 MHz
SPAN 5·000 kHz

Fig.15

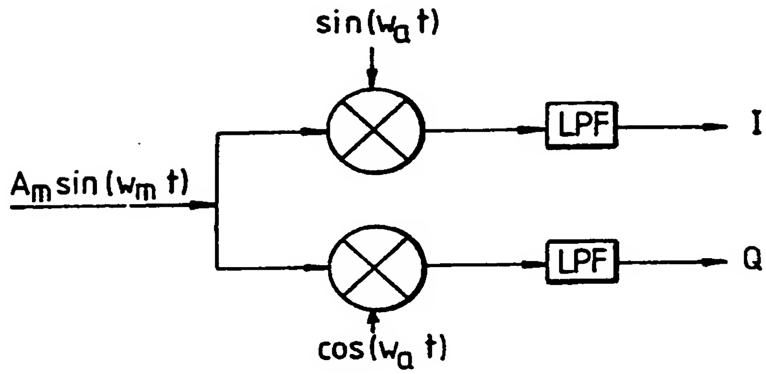
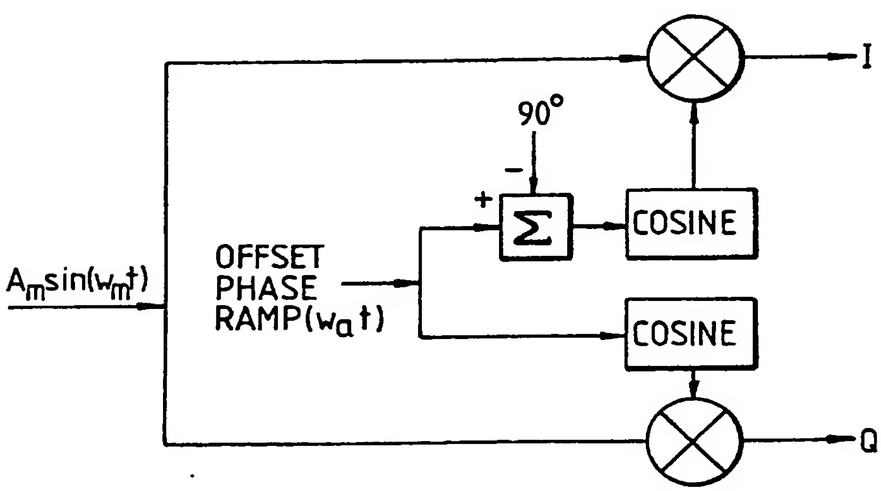

sin(wa t)
Am sin (wm t)
LPF
I
LPF
Q
cos(wa t)

Fig.16


I
90°
−
+
Σ
COSINE
Am sin(wm t)
OFFSET
PHASE
RAMP(wa t)
COSINE
Q

CORRECTION OF ERRORS IN A ZERO-IF TRANSMITTER

This invention relates to a method and means for the correction of errors present in a zero-IF transmitter, e,g, in a radio transmitter.

The principle of the zero-IF, or direct conversion, transceiver is well known and products are now on the market employing this technology. This type of transceiver has the advantage of avoiding the image problems inherent in conventional superheterodyne designs by using a local oscillator that is in tune with the rf carrier in order to use an intermediate frequency of dc. This technique relies on the fact that the consequent 'aliasing' of the baseband signal can be unfolded using digital signal processing (DSP) if a quadrature mixing arrangement as shown in Figure 1 is employed with the local oscillator, as disclosed in British patent No. 2106734.

However, one of the major problems inherent in the design of direct conversion transceivers is that of obtaining accurately balanced quadrature channels. The signal processing relies on the fact that these two channels must have identical amplitude characteristics, together with exact phase quadrature between the local oscillator inputs to the mixers, and there can be no local oscillator (LO) feedthrough. The penalty if these conditions are not met is that, in receive mode, the sensitivity of the demodulator is reduced due to the addition of dc terms through self-mixing produced by the LO feedthrough, while the amplitude and phase errors cause intermodulation products of the modulating signal

and the offset between the carrier and the LO to break through. In transmit mode, however, the effects of these three types of error are subtly different. Now, the LO feedthrough serves to provide an additional unwanted component at the carrier/LO frequency, while the amplitude and phase errors combine to produce in-band interference due to inadequate suppression of spectrally reversed sidebands as shown in Figure 2 for the case of USB SSB.

In practice, however, the acceptable performance limits on the transceiver, determined by auditory quality in receive and spurious signal suppression in transmit mode, mean that the amplitude phase, and feedthrough requirements do not have to be met exactly, but a small degree of mismatch is acceptable. Unfortunately, even with this reduction in specification, current mixer and analogue filter designs still cannot satisfy the channel matching requirements, due to component tolerances.British patent application No. 8505923 shows that in the receive mode it is possible to implement real-time error correction routines that operate on the digitised quadrature channels within the tranceiver's DSP chip, to provide the necessary demodulation performance. In the receive case this is a relatively straightforward task, since the erroneous channel signals are at baseband frequencies, which are easily processed by the DSP chip. However, for the transmit mode, the full effects of the filter and mixer errors are not observed until the channel signals are mixed up to radio frequencies, and thus techniques need to be found to correct imbalances in high frequency quadrature channels.

In practice, the difficulty lies in measuring the errors in order to discover the magnitude of the corrections necessary, rather than in actually applying

these correction factors. However, before describing the full error correction scheme of the present invention, it is helpful to explain the feedforward techniques by which digital processing at baseband is able to correct the RF errors.

1)Channel Amplitude Imbalance

At the rf combiner, the two channels will have different maximum amplitudes, due to the small mismatches between the filters and mixers in the signal path. But, this imbalance is easily corrected by applying a relevant scaling factor to each of the quadrature based signals before the baseband quadrature signals are output to the D/A converters.

2) Deviation from Phase Quadrature

The analogue imperfections between the filter and mixer pairs also give rise to unequal phase shifts in the two channels. Practically, this can be treated as a lumped effect of a deviation from phase quadrature between the two local oscillator feeds to the mixers, as shown in Figure 3. In order to correct this error, a phase shift, equal in magnitude to this deviation from quadrature, is applied to the in-phase baseband signal. However, this phase shift needs to be constant over the whole bandwidth of the baseband signal, else the correction would not work. In order to achieve such an effect, the correction factor may be applied in the manner shown in Figure 4 for the various modulation formats. The mathematical proofs of these correction schemes are given in Appendix A.

3) Local Oscillator Feedthrough

All practical RF mixers have limited LO-RF

isolation due to the nature of their design and manufacture. Unfortunately, the isolation provided is not good enough for the direct conversion radio, and thus must be improved. The effect of this LO feedthrough upon transmission is exactly the same as that created by any unwanted dc bias at the IF port of the mixer, in that the component at the LO frequency is radiated at a different level than that which the modulation requires, and thus the received signal is erroneous.

Consequently, in order to correct for the poor isolation, it is possible to add a specific dc bias, at the IF port of a mixer, that serves to reduce the unwanted LO frequency present at the RF port. The results of the addition of such a signal are shown very clearly in Figure 5 for a single mixer at various LO frequencies. It is noted that the effectiveness of this technique is dependent upon the frequency of operation, and this is because, as the frequency changes, so the phase shift in the LO-RF leakage path alters. However, the dc bias can only completely cancel the unwanted LO component when their phase relationship is true antiphase at the RF port, although marked improvement of the isolation is seen at every frequency.

To summarise, the amplitude imbalance is corrected by appropriately scaling the two quadrature baseband signals, the phase imbalance is compensated for by applying a single correction phase shift in the baseband I channel, and the local oscillator feedthrough can be largely eliminated by applying an appropriate polarity dc bias to each of the baseband channels just prior to the D/A converters.

Returning to the question of measuring the three types of imbalance at RF frequencies in order to determine the magnitude of the correction factors

necessary, the most straightforward scheme utilises a feedback arrangement to monitor the result of applying these factors. Thus the correction factors are found directly. There are several methods available for determining the effect of the baseband corrections described above:

i ) The use of a very fast A/D converter and signal processing chip, in order to be able to digitally analyse and process the channel signals to monitor and drive the correction of errors.

ii ) The introduction of an additional signal path in the radio, using one extra mixer, to enable real-time monitoring and correction of the transmitted signal.

iii) The introduction of an additional signal path in the radio, using two extra quadrature mixers, to enable real-time monitoring and correction of the transmitted signal.

These techniques are able to provide a real-time automatic error correction scheme, which is desirable, but, unfortunately, each of these methods have inherent drawbacks.

Method i) relies upon the availability of extremely fast A/D converters, not to mention the fact that general purpose DSP chips are not currently capable of running at sample rates fast enough to process HF signals. Considering, for example, the TMS 320C25 microprocessor, which runs at a nominal instruction rate of 100 ns, then the implementation of a complete radio would allow a sampling rates of only 100 kHz, and, even then, the processing load requires that four TMS chips would be

necessary.

Method ii) suffers from the fact that a single mixer in this extra monitoring path can only provide one signal that contains the effects of all the errors, and thus, unlike the receive case, there is not the possibility of comparison between the two channels to determine the errors. In addition, the single signal has the disadvantage that the resultant overlaying of the image and wanted signals, together with the random nature of the modulating signal, serves to confuse the effects of the errors to such an extent that correction becomes an almost impossible task.

Method iii) has the major drawback that introducing quadrature mixing into the monitoring path also introduces extra errors into the radio, even though they are not in the actual transmission path. Thus the problem of balancing the channels in the monitor path is also introduced, and if the balance is not achieved, the transmit error corrections cannot be implemented.

According to the present invention there is provided a method for the correction of errors in a zero -IF transmitter using SSB mode, comprising the step of successively

generating an auxiliary local oscillator signal the frequency of which is offset from the transmitter rf oscillator,

utilising the auxiliary oscillator signal to effect reduction of rf oscillator feedthrough,

effecting balancing of the quadrature channel amplitudes, and

utilising the auxiliary oscialltor signal to effect reduction of phase deviation from quadrature in

the quadrature channels.

The above method requires the introduction of an additional signal path in the radio, using one extra mixer, to enable a calibration exercise to measure the values of correction factors for the channel imbalances, which are then used in a real-time static correction mechanism. This method however, does not suffer from the same disadvantage as method ii) above, even though they are based upon the same principle architecture. The reason for this is that an appropriately simple modulating system may be chosen as the calibration test signal, such that the overlaying of the interfering signals upon the wanted channel does not hinder, but infact aids the measurement process.

The method of the invention is based around a set of recursive algorithms that apply trial baseband correction factors and measure their effect, using the result to drive the correction factors to their exact values, and consequently eliminate the effects of the RF error mechanisms. In principle, this calibration exercise can be performed using any known modulating signal within the audio baseband frequency range from 300 to 3000 Hz. However, for ease of explanation, a modulating tone at 650 Hz will be assumed.

Embodiments of the invention will be described with reference to Figures 6 - 16 of the accompanying drawings wherein

Figure 6 is a schematic diagram of the radio architecture necessary for the transmission error correction calibration routine of the invention,

Figure 7 illustrates the spectral output of an upper sideband SSB modulator with certain errors,

Figure 8 illustrates the spectrum resulting from the mixing down the imperfect signal of Figure 7 with an auxiliary local oscillator offset,

Figure 9 shows a local oscillator feedthrough correction algorithm,

Figure 10 shows an algorithm for the implementation of amplitude balance correction,

Figure 11 shows an algorithm for the correction of phase quadrature deviation,

Figure 12 is a block diagram of a calibration routine for transmission error correction,

Figure 13 shows an application algorithm for local oscillator feedthrough correction,

Figure 14 shows transmitted spectra for USB SSB modulation of a 1kHz tone without and with error correction,

Figure 15 illustrates generation of baseband signals for the Weaver method of SSB modulation, and

Figure 16 illustrates SSB baseband signal generation within the DSP chip.

The architecture used for the error calibration method is shown in Figure 6, and it is noted that three extra signal paths back to the DSP chip have been included. There are two balanced peak detector circuits 60,61 that sense the long term maximum voltages on the two quadrature channels just before they are summed in the RF combiner 62, and these levels are then passed into the DSP chip 63 through an A/D converter 64. The third additional path 65 is a monitor path for the generated signal, and is only connected to the combiner output via switch 66 whilst calibration is in progress. (In normal transmission mode, the combiner output is passed directly to the radio front-end). This monitor path 65 incorporates a further mixer 67 to convert the RF signal back to baseband without introducing folding around dc. This is achieved by using a local oscillator frequency that is offset from the quadrature LO frequency by a small amount, e.g. an audio frequency. After low-pass filtering 68, this baseband signal enters the DSP chip via a second A/D converter 69.

The reason for the two peak detector circuits 60,61 is that the amplitude and phase errors produce the same effect, and thus it is useful to provide an independent means for balancing the channel amplitudes. Furthermore, there is the added advantage that, although most of the correction is performed by calibration, the amplitude balancing is able to be run continuously.

In order to perform the calibration, the transmitter is run in upper sideband SSB mode, which allows the error effects to be seen unambiguously. If, for example, angle modulation were used, the single sinusoid used for calibration would generate many sidebands due to the non-linear process, and it would be impossible to extract information about the errors. Considering the USB SSB modulation of the 650 Hz sinewave using, for example, the Weaver method, and assuming all correction factors are set to zero, then an analysis can be performed on the effect of the errors. For an amplitude imbalance of 10%, a phase deviation from quadrature of 3 degrees, and a local oscillator isolation of 40dB, then using a local oscillator at 50 MHz and +17dBm, an offset frequency of 1650 Hz, and with the channel signal level at +5 dBm, the output of the combiner is given by

$$S = Aw.\sin(W.t) + Al.\sin(Wl.t) + Au.\sin(Wu.t) \qquad (1)$$

where W is 49.999 MHz and is the wanted signal, Wl, is 50.0 MHZ and is the local oscillator feedthrough, and the unwanted signal, Wu, is 50.001 MHz. The relative signal levels are as shown in Figure 7, and upon mixing down with an auxiliary local oscillator at 50.002 MHz, followed by simple low pass filtering, the three separate tones are preserved at baseband, albeit in reversed spectral order, as shown in Figure 8.

From the above theory, it is clearly seen that the calibration architecture allows measurement of the magnitude of the unwanted signals due to the rf errors. Within the DSP chip two bandpass filters (70, 71, Figure 6) are used to separate the two mixed-down error signals, and the two signals from the peak detectors 60, 61 are also entered into the DSP chip. Consequently, four signals are available to determine the performance of the system.

In order to eliminate the error effects, it is vital that the errors are corrected in the right order. This is because the amplitude and phase effects are identical, and the LO feedthrough impacts upon the performance of the amplitude correction scheme. Consequently, it is first necessary to improve the LO rf isolation as much as possible.

Unfortunately, there is only one error signal (at 2 kHz) for this correction, whereas there are two independent feedthroughs to eliminate. As a result, a dual recursive scheme is necessary, whereby the error signal is fist minimised by considering only the I channel, and then the Q channel is corrected to further reduce the energy at 2 kHz. If the improvement in performance is still not adequate, then the correction procedure described above must be repeated until the desired level of performance is achieved. In practical tests, however, it has been found that the feedthrough can be eliminated independently, thus removing the need for repeated application of the correction algorithm, as given in Figure 9.

Once the LO feedthrough has been removed,the channel amplitudes may be balanced. This is a relatively simple procedure, requiring that the ratio of the two

peak detected signals is found. This determines the amount of the imbalance, and the larger of the two channels is then reduced by the scale factor just calculated. The algorithm for this procedure is given in Figure 10.

This leaves only the phase deviation from quadrature to be removed. In order to achieve this, a recursive routine is implemented where trial values of the correction factor d are used until the remaining error signal at 1kHz is reduced to zero. Consequently, the value of the correction factor (valid for all the modulation formats) has been determined. The algorithm used is shown in Figure 11.

Up to this point, the calibration has been performed for one particular value of local oscillator frequency, and the correction coefficients obtained must now be stored in RAM for later retrieval under operational conditions. The calibration exercise described above is then repeated for as many LO frequencies as is deemed necessary for the radio. Once this completed calibration, as detailed in Figure 12, is finished, the radio is then operational, and the chosen transmission band and mode determine the correction coefficients to be applied in the appropriate manner as given in Figures 10, 4, and 13. Furthermore, the radio is also able to continuously update the amplitude scaling factor, as the peak detector circuits, and their paths back to the DSP chip, are a permanent feature of the radio.

In conclusion, this invention provides a method for correcting the RF errors that manifest themselves upon transmission in a zero-IF radio, and which are due to the quadrature architecture that is necessary in this type of transceiver. This method is based upon baseband

correction mechanisms that are designed to compensate in a feedforward manner, for the three types of RF error, namely the channel amplitude imbalance, the phase deviation from quadrature, and the local oscillator feedthrough. The scheme itself is a calibration procedure that is equally applicable to true direct conversion radios, or to conventional superheterodyne radios using a final IF at dc. The scheme proposed has been demonstrated to work in practice, and the benefit obtained is clearly illustrated in Figure 14, where the comparison between an uncorrected and a corrected zero-IF transmitter is shown for USB SSB modulation of a 1kHz tone.

## APPENDIX A

### Mathematical Basis for the Phase Correction Techniques

The methods for applying the phase correction factors are detailed in Figure 4 for each of the modulation modes, except for AM which is generated using only one of the two quadrature channels, and thus does not require phase compensation. This appendix describes the derivation of these techniques, based upon the representation of the phase imbalance as a lumped phase deviation from quadrature, as shown in Figure 3.

### Angle Modulation.

Considering an FM or PM modular implementation, the baseband I and Q channels are given by

$$I = \cos[V(t)] \tag{A.1a}$$
$$Q = \sin[V(t)] \tag{A.1b}$$

where V(t) is generated from the modulating signal.

After mixing with perfect quadrature LO signals and summing the resultant mixer outputs, then, assuming perfect channel balance, the modulated signal is given by

$$S = 2C.\sin[w_o t+V(t)] \tag{A.2}$$

whereas, mixing the I and Q signals of equation (A.1) with the LO signals of Figure 3 gives

$$S = C.** \sin[w_o t+V(t)] + \sin[w_o t+V(t)+d] - \sin[w_o t-V(t)+d]* \tag{A.3}$$

where C is a constant equal to the gain through the filter and mixer in each channel and it is assumed that the amplitude correction, as detailed earlier, has been used to balance the two channels. It is clearly seen from equation (A.3) that an unwanted image signal at $w_o t-v(t)$ is produced by the phase quadrature deviation d.

In order to compensate for the quadrature deviation, and remove the image signal, a phase shift of -d is introduced into the I channel, such that it becomes sin[V(t)-d]. Mixing with the local oscillator signals given in Figure 3 gives

$$I' = C.** \sin[w_o t+V(t)-d] + \sin[w_o t-V(t)+d]* \tag{A.4a}$$

$$Q' = C.** \sin[w_o t+V(t)+d] - \sin[w_o t-V(t)+d]* \tag{A.4b}$$

and upon summing the two channels, and assuming that the channel amplitudes have been equalised, the modulated signal is now given by

$$S = C.** \sin[w_o t+V(t)-d] + \sin[w_o t+V(t)+d]* \tag{A.5}$$

Expansion of the sinusoids leads to

$$S = 2C.\sin[w_o t+V(t)].\cos(d) \qquad (A.6)$$

and comparison of equation (A.6) with equations (A.3) and (A.2) shows that the effect of introducing an equal and opposite phase shift in the I channel is to eliminate the image response caused by LO quadrature deviation, d, at the expense of reducing the amplitude of the wanted signal by the factor cos(d).

In practice the value of d should lie in the range $-3^o$**d**$3^o$, and thus the maximum reduction in signal amplitude is 0.14%, which is a negligible effect when compared with the distortion produced without any phase error correction. The method of insertion of the compensating factor d, in the I channel is shown in Figure 4(a), and is based upon the fact that the I and Q baseband signals for FM and PM are generated by using a look-up table (LUT) for the cosine function, with the sine values being determined by subtracting 90 degrees before accessing the LUT.

This works on the principle that once the modulating signal has been digitised (if it is analogue modulation rather than data being transmitted), then V(t) is a number representing the magnitude of [the integral of] the modulating signal for [FM] PM. This number given by V(t) is also a value of phase in the range -180 to +180 degrees. Consequently, adding in a fixed offset such as -90 degrees, or-d degrees, is a simple matter of adding in the number representing the phase offset in the scaled range available in the DSP chip. Furthermore, this method of inserting the phase correction factor neatly manages to apply a constant phase shift over the whole of the modulating signal bandwidth - which stretches to 12.5 kHz for wideband FM.

Single Sideband

SSB requires a rather more complicated correction scheme than either of the angle modulations, since the final baseband signals are not obtained by simply accessing a look-up table. Rather the modulating signal is mixed with quadrature offset frequency sinusoids, as shown in Figure 15. However, these offset frequency sinusoids are generated from look-up tables, as shown in Figure 16, and thus the necessary phase shift can be introduced into these sinusoids in order to correct the mixer induced phase errors.

Considering first the case of upper sideband SSB with no errors, the I and Q channels at the output of the DSP chip are given by

$$I = Am.\cos[(w_m - w_a)t] \qquad (A.7a)$$
$$Q = Am.\sin[(w_m - w_a)t] \qquad (A.7b)$$

and mixing these with perfect quadrature LO signals, followed by addition of the two channels, gives the desired USB signal

$$S = C.Am.\sin(Wt) \qquad (A.8)$$

where $W = w_o + w_m - w_a$.

Considering now the correction of the phase deviation from quadrature, introducing d into the in-phase offset sinusoid, such that it becomes $\sin[w_a t + d]$, as shown in Figure 4(b), leads to I and Q channel outputs from the DSP chip of

$$I = Am.\cos[(w_m - w_a)t - d]$$
(A.9a)
$$Q = Am.\sin[(w_m - w_a)t]$$

(A.9b)

Mixing these terms up with the LO signals shown in Figure 3 gives

$$I' = C.Am^{**} \sin[(w_o+w_m-w_a)t-d] + \sin[(w_o-w_m+w_a)t+d]^{**} \quad \text{(A.10a)}$$

$$Q' = C.Am^{**} \sin[w_o+w_m-w_a)t+d] - \sin[w_o-w_m+w_a)t=d]^{**} \quad \text{(A.10b)}$$

and addition of I' and Q' produces

$$S = C.Am^{**} \sin(W-d) + \sin(W+d)^{**} \quad \text{(A.11)}$$

Expansion of these sinusoids give the result

$$S = 2C.Am.\sin(W).\cos(d) \quad \text{(A.12)}$$

and, once again, it is seen that the introduction of a compensating phase shift at baseband has given rise to the desired modulated signal, with the consequent reduction in output amplitude.

Now, considering the case of lower sideband SSB, the desired output signal obtained with no errors is

$$S = 2C.Am.\sin(w't) \quad \text{(A.13)}$$

where $W' = w_o-w_m+w_a$. When the error correction as detailed in Figure 4(c) is applied, the baseband I and Q channels become

$$I = Am.\cos[(w_m-w_a)t+d] \quad \text{(A.14a)}$$
$$Q = Am.\sin[(w_m-w_a)t] \quad \text{(A.14b)}$$

and, mixing with the erroneous local oscillator signals

of Figure 3 gives

$$I' = C.Am^{**} \sin[(w_o+w_m-w_a)t+d] + \sin[(w_o-w_m+w_a)t+d] \quad (A.15a)$$

$$Q' = C.Am^{**} \sin[(w_o+w_m-w_a)t+d)^{**} \sin[(w_o-w_m+w_a)t+d]^{**} \quad (A.15b)$$

Subtracting the Q from the I channel, the output signal becomes

$$S = C.Am^{**} \sin(W't-d) + \sin(W't+d)^{**} \quad (A.16)$$

and expanding the sinusoids produces the result

$$S = 2C.Am.\sin(W't).\cos(d) \quad (A.17)$$

from which it is apparent that the inclusion of the compensating factor as shown in Figure 4(c) has indeed corrected for the phase shift imbalance between the two quadrature channels.

To summarise, the phase correction is performed by adding a phase correction factor into the baseband in-phase channel. This correction factor has the magnitude of the quadrature deviation at the LO, and its sign is dependent upon the modulation mode of operation.

CLAIMS

1. A method for the correction of errors in a zero -IF transmitter using SSB mode, comprising the step of successively

generating an auxiliary local oscillator signal the frequency of which is offset from the transmitter rf oscillator,

utilising the auxiliary oscillator signal to effect reduction of rf oscillator feedthrough,

effecting balancing of the quadrature channel amplitudes, and

utilising the auxiliary oscialltor signal to effect reduction of phase deviation from quadrature in the quadrature channels.

2. A method according to Claim 1 including the step of mixing the transmitter output with the auxiliary oscillator signal to form an error signal for feedback for the reduction in rf oscillator feedthrough and the reduction in phase deviation.

3. A method for the correction of errors in a zero -IF transmitter using SSB mode, wherein digital baseband signals produced in quadrature channels in a digital signal processing (DSP) arrangement are separately converted into analogue signals in each channel prior to be mixed with an rf oscillator signal in quadrature, the resulting quadrature rf signals being summed to from the transmittrr output or IF signal, the method comprising the steps in succession of

generating an auxiliary local oscillator signal the frequency of which is offset by a small amount from the frequency of the rf oscillator,

mixing the auxiliary oscillator signal with the transmitter output,

low pass filtering the resultant mixed signal to form an analogue error signal,

converting the analogue signal to a digital

signal,

first deriving from said digital signal a digital error at the auxiliary oscillator offset frequency and feeding said first derived digital error signal to the DSP arrangement to reduce the local rf oscillator feedthrough,

subsequently detecting the peak amplitudes of the quadrature rf signals,

low pass filtering and converting said peak amplitudes into separate digital signals and feeding said separate digital signals to the DSP arrangement to effect balancing of the quadrature channel amplitudes, and

finally second deriving from said digital signal a signal at a frequency other than the auxiliary oscillator offset frequency and feeding said second derived digital signal to the DSP arrangement to effect reduction of the phase deviation from quadrature in the quadrature channels.

4. A method according to claim 3 wherein said auxiliary oscillator frequency offset is an audio frequency.

5. A method according to claim 3 or 4 wherein said first determined digital signal is obtained by band-pass filtering of the digital error signal.

6. A method according to claim 3, 4 or 5 wherein the second determined digital signal is obtained by band-pass filtering of the digital error signal.

7. A method according to claim 6 wherein the second determined digital signal is obtained by band-pass filtering of the digital error signal at half the offset frequency.

8. A method according to claim 3 wherein the reduction of the phase deviation from quadrature channels is effected by addition of a compensating phase shift into the auxiliary local oscillator signal fed to one of the quadrature channels prior to mixing of the

auxiliary oscillator signals with the transmitter output.

9. A method for the correction of errors in a digital zero-IF transmitter using SSB mode, wherein digital baseband signals produced in quadrature channels in a digital signal processing (DSP) arrangement are separately converted into analogue signals in each channel prior to being mixed with a local rf oscillator signal and then summed to form the transmitter output or IF signal, the arrangement comprising

an auxiliary local oscillator for generating a signal the frequency of which is offset by a small amount from the frequency of the rf oscillator,

means for mixing the auxiliary oscillator signal with the transmitter output,

means for low pass filtering the resultant mixed signal to form an error signal

means for converting the analogue error signal to a digital signal,

first means for band pass filtering said digital signal at the auxiliary oscillator offset frequency and feeding the band pass filtered signal to the DSP arrangement to reduce the rf local oscillator feedthrough,

means for detecting the peak amplitudes of the quadrature rf signals,

means for low pass filtering and converting said peak amplitudes into separate digital signals and means for feeding said separate digital signals to the DSP arrangement to effect balancing of the quadrature channel amplitudes, and

second means for band pass filtering the digital error signal at a frequency other than the auxiliary oscillator offset frequency and feeding the band pass filtered digital signal to the DSP arrangement to effect reduction of the phase deviation from quadrature in the quadrature channels.

10. An arrangement according to claim 9 wherein the second band-pass filtering means filters the digital error signal at half the offset frequency.

11. An arrangement according to claim 9 or 10 including means for adding to the auxiliary oscillator signal fed to one of the quadrature channels a compensating phase shift to effect the reduction of the phase deviation from quadrature in the quadrature channel.

12. An arrangement for the correction of errors in a zero-IF transmitter operated in SSB mode substantially as described with reference to Figure 6 of the accompanying drawings.

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